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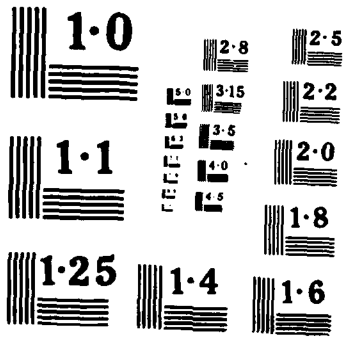
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DIELECTRIC ANTENNAS FOR MILLIMETER-WAVE APPLICATIONS

TECHNICAL REPORT

G. M. WILKINS
R. MITTRA

MAY 1986

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20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

In this report, a number of designs of dielectric antennas for millimeter-wave applications are presented. Included are details of the basic configuration and performance of these devices, along with discussions of modifications and their use as components in millimeter-wave integrated circuits.

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TABLE OF CONTENTS

LIST OF FIGURES

I. INTRODUCTION

With the development of dielectric guiding media and integrated circuits which operate in the millimeter-wave region, there arises the need for additional dielectric components. A lot of interest has been shown for one such component, the dielectric antenna, since it eliminates the problems associated with feeding and matching such structures for efficient power transfer. This type of antenna is also of interest because of its relatively simple design, compactness, light weight, and low cost. It is capable of electronically scanning the radiation beam by varying the operating frequency and is, therefore, capable of replacing slow, mechanically scanned structures. Furthermore, its relatively high resolution, ability to operate at night, penetrability through dust and fog, and several other desirable characteristics make this antenna a prime candidate for military applications, where all of these factors must be taken into account.

This report is a review of some of the literature written on this component over the last few years. It points out some problems the antenna encounters, highlights its numerous capabilities, and describes some of the attempts to improve the design of this antenna.

II. DIELECTRIC ROD ANTENNAS

The simplest configuration of the dielectric antenna, the dielectric rod antenna, is examined extensively by Kobayashi [1] and somewhat by Paleta [2]. This antenna is a surface-wave antenna, i.e., the wave travels along the outer surface of the dielectric. It is a good directional radiator in the endfire direction, although somewhat limited due to relatively low gain. Figure 1 is the schematic

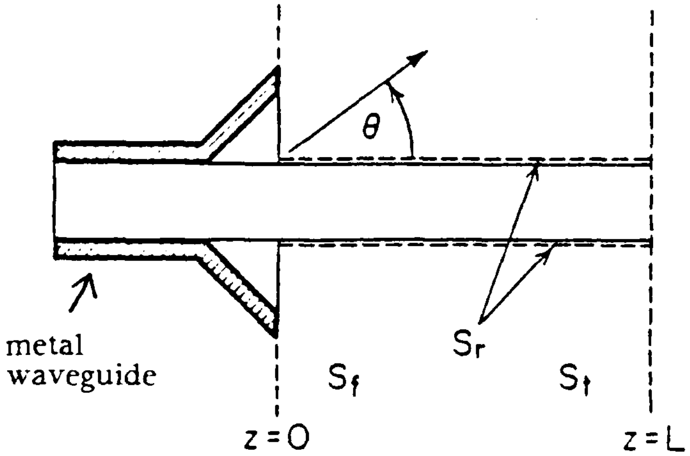


Figure 1. Schematic diagram of dielectric rod antennas.

diagram for the dielectric rod antenna. The radiation takes place at the discontinuities, mainly the feed point and the terminal point. The radiation pattern, $F(\theta)$, due mostly to terminal radiation, is calculated by integration over the terminal aperture S_t, and is approximately given by $F(\theta) \approx 1/\psi$, where $\psi = (k_o L/2)(\cos\theta - r)$, $r = \beta_z / k_o$, k_o is the free-space wave number, L is the length of the rod, and β_z is the dielectric phase constant in the z-direction. There are no side lobes associated with this pattern. Taking the feed point into account results in side lobes which tend to sharpen the main lobe. This serves as the motivation for the experimental determination of a design scheme for these antennas, as discussed by Kobayashi.

Figure 2 shows the basic configuration for maximum gain antennas, in which the feed taper establishes a surface wave along the straight section and reduces loss at the feed point, and the terminal taper reduces reflection and radiation which are caused by the abrupt discontinuity at the terminal end. Experiment has shown the usefulness of this design scheme, but it has also given rise to several other problems which must be considered. Some of them are as follows:

1) In order to determine the length of the rod, the propagation constant along the rod must be known. No definite method of calculation is available for rods with small rectangular cross sections. McLevige [3], Paleta [2], Yang [4], Rudokas [5], and Bhooshan [6] all discuss this problem, but there still exists the problem of the taper.

2) Even if the propagation constant along a straight rod is calculated, it is difficult to estimate the propagation constant along the taper.

3) The leaky-wave radiation from the feed taper and terminal taper must be taken into account.

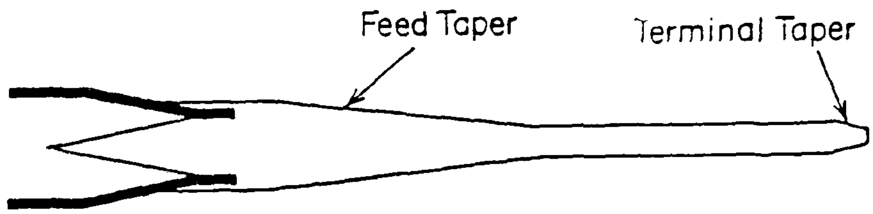


Figure 2. Basic configuration of maximum gain antennas.

4) In addition to the leaky-wave radiation from tapers, another factor in determining the radiation pattern is the radiation from the abrupt discontinuity at the junction between the metal waveguide and dielectric rod.

5) Tapering is not easy to accomplish, especially for small rectangular cross sections. The structure may also lack mechanical strength due to this configuration.

Table 1 is a list of the factors which must be considered in the design of a dielectric rod antenna, the factors being illustrated in Figure 3. Kobayashi takes $A_m = A_r$, $B_m = B_r$, and does not present precise design principles, but instead gives an overview of the experimental results which may be helpful in the actual design of low side-lobe pattern antennas.

Kobayashi discusses the overall performance of several antennas by measuring the far-field pattern. The dielectric material used is polyethylene ($\epsilon_r = 2.33$) with a cross section of 3.1 mm x 1.55 mm. In general, the materials used have dielectric constants on the order of 2-25. The frequency of testing is 81.5 GHz. Experiments show that tapering the transition into the metal waveguide in the x-z plane (H-plane) produces better results than tapering in the y-z plane (E-plane), and better matching and mechanical support are obtained by tapering in just the H-plane. For simplicity, a linear taper is used. He finds experimentally that return loss is minimal for L_t = 10 mm. The performance of these antennas is discussed below.

The simplest configuration of the dielectric rod antenna is the straight rod antenna (Figure 4 (a)). The radiation takes place at only the feed end and the terminal end, and may therefore be regarded as a two-element array with an element pattern equal to that of a rectangular aperture. This interferes with the desired

TABLE 1

DESIGN FACTORS FOR A DIELECTRIC ROD ANTENNA

Common Factors:

Frequency range
Metal Waveguide (cross-sectional dimensions)
Material of the rod (dielectric constant, loss tangent, etc.)

Transition Factors:

Length of transition
Taper of the metal waveguide
Taper of the rod

Feed Factors:

Length of feed
Flare angle of the horn
Taper of the rod

Main Antenna Factors:

Length of main antenna
Taper of the rod (including the terminal taper)

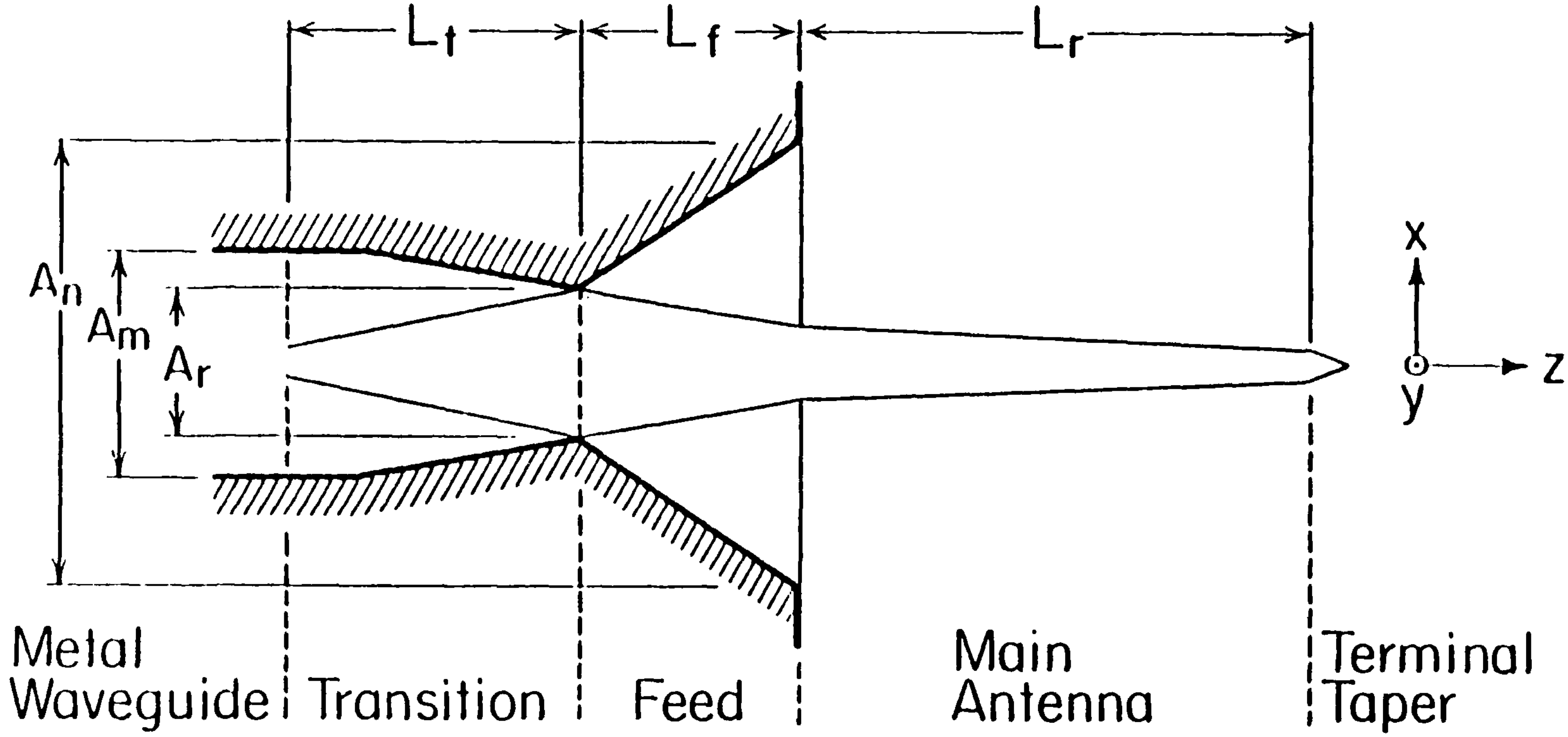


Figure 3. Dimensions of a dielectric rod antenna in xz-plane. In yz-plane, B_n, B_m and B_r are used instead of A_n, A_m and A_r.

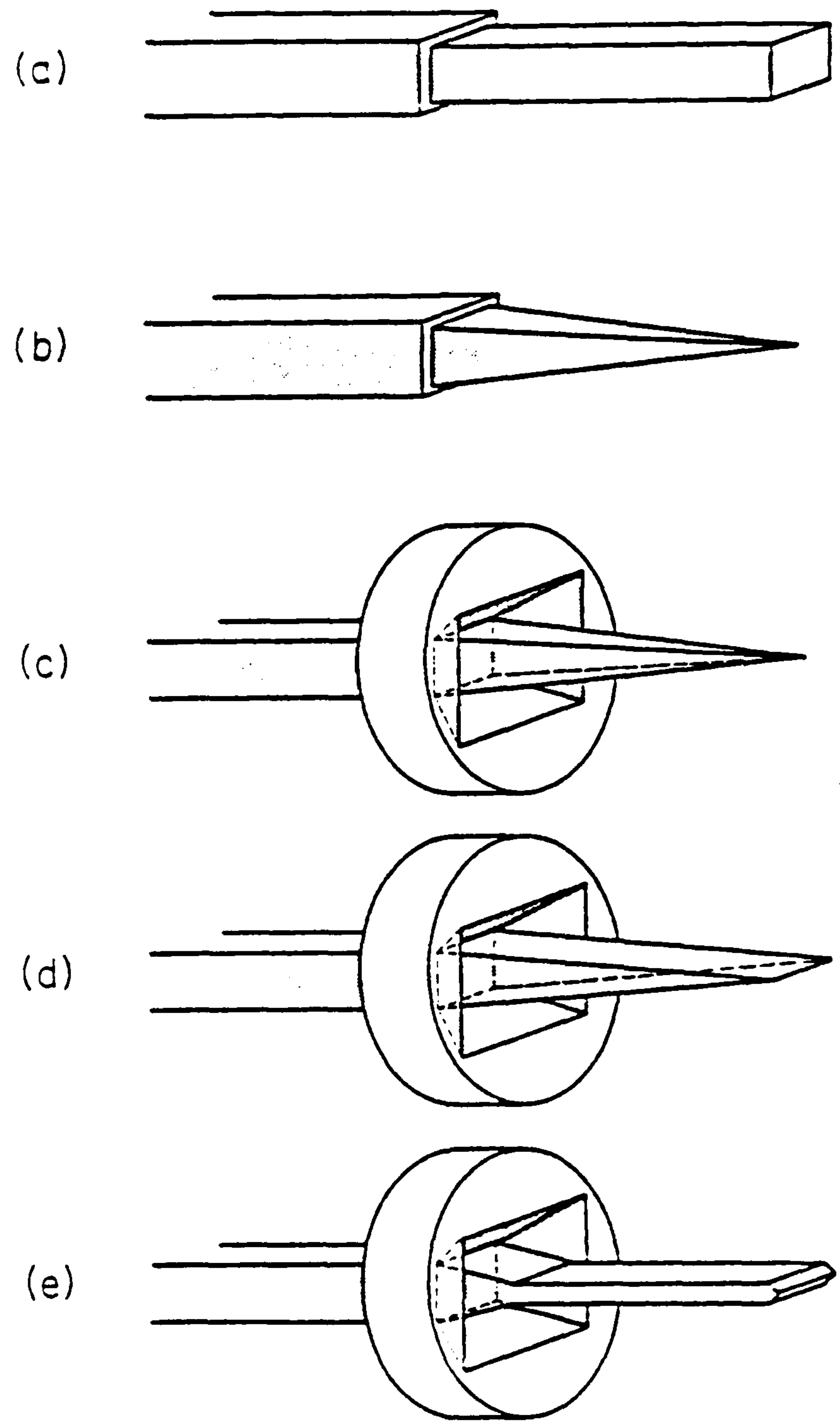


Figure 4. Examples of dielectric rod antennas with rectangular cross-section.

endfire radiation. If $L_r >> \lambda$, many lobes will be seen, whose envelope is the patterns of the waveguide apertures.

The tapered rod antenna (Figure 4 (b)) is an improvement of the straight rod antenna since it reduces radiation and reflection from the abrupt discontinuity at the terminal end. But leaky-wave radiation takes place from the taper in addition to the surface-wave radiation from the discontinuity at the feed end, thus producing distortions in the main lobe as a result of interference from surface-wave, feed-end, and terminal-end radiations. Also, the feed-end radiation produces relatively high side-lobe levels. Furthermore, the taper is neither strong nor easy to make.

A solution to the problem of feed-end radiation is the addition of a feed horn (Figure 4 (c)). This reduces the feed-end radiation and increases the efficiency of launching the surface-wave onto the rod, as well as reducing the side-lobe level. Unfortunately, it also reduces gain and increases the HPBW. If the length is increased, the gain will be increased, but so will the side-lobe level. As before, the taper is neither strong nor easy to make.

The E-plane tapered rod antenna with feed horn (Figure 4 (d)) improves the mechanical situation with no sacrifice in antenna characteristics and sometimes with a little improvement. In addition, Kobayashi tries an H-plane taper and a both-plane taper, and finds that neither performs as well as the E-plane taper, and the performances of the H-plane taper is always worse than that for both-plane taper.

Kobayashi further investigates the dielectric rod antenna by altering the last configuration in accordance with Zucker's design principles ([1] Appendix, Figure 4(e)), and by increasing the dielectric constant ϵ_r. In addition to the unsatisfactory performance of this latter set of antennas (low gain, high side-lobe levels, very distorted main lobe), just determining the dielectric constants at these frequencies may present a problem due to the elaborate equipment requirements, large sample requirements, and low wave-number resolution. It is more convenient to use a material whose dielectric constant is already known. Although Ahn [7] describes a simple technique which overcomes these difficulties, Zucker's design principle discourages the use of higher dielectric constant material since the principle states that the higher the gain, the thinner the rod, which somewhat cancels the physical strength of the higher dielectric constant material.

While Kobayashi investigates considerably the effects of altering the shape of the rod taper, Chang and Mittra [8] are more concerned with the transition at the waveguide-dielectric interface. At millimeter-wave frequencies, there exists the need to devise a transition which is as short as possible and with an acceptably small insertion loss. Chang and Mittra conclude that regardless of the type and length of the transition from metal waveguide to the antenna, it is possible to obtain almost identical gain radiation patterns in the frequency range 76-80 GHz provided that the parameters L_a and L_w are kept constant (see Figure 5). Not only is a smooth transition not needed, but it will suffice to have just enough L_t to attach the dielectric antenna to the waveguide so that it will adhere to it. Also, as long as $L_a - L_w \approx 1.0\lambda - 1.5\lambda$, the gain characteristic is good enough to match the patterns obtained from well-made symmetric antennas. Therefore, in this frequency range a specific transition shape is not needed to obtain a reasonable performance from this antenna.

Doran and Mittra [9] also perform experiments on the dielectric rod antenna. Their experiments are geared toward finding an optimum shape for the rod outside of the antenna. In their testing they find that at 81 and 220 GHz, of the antennas shown is Figure 6, the wedge antenna has the highest gain of all the antennas

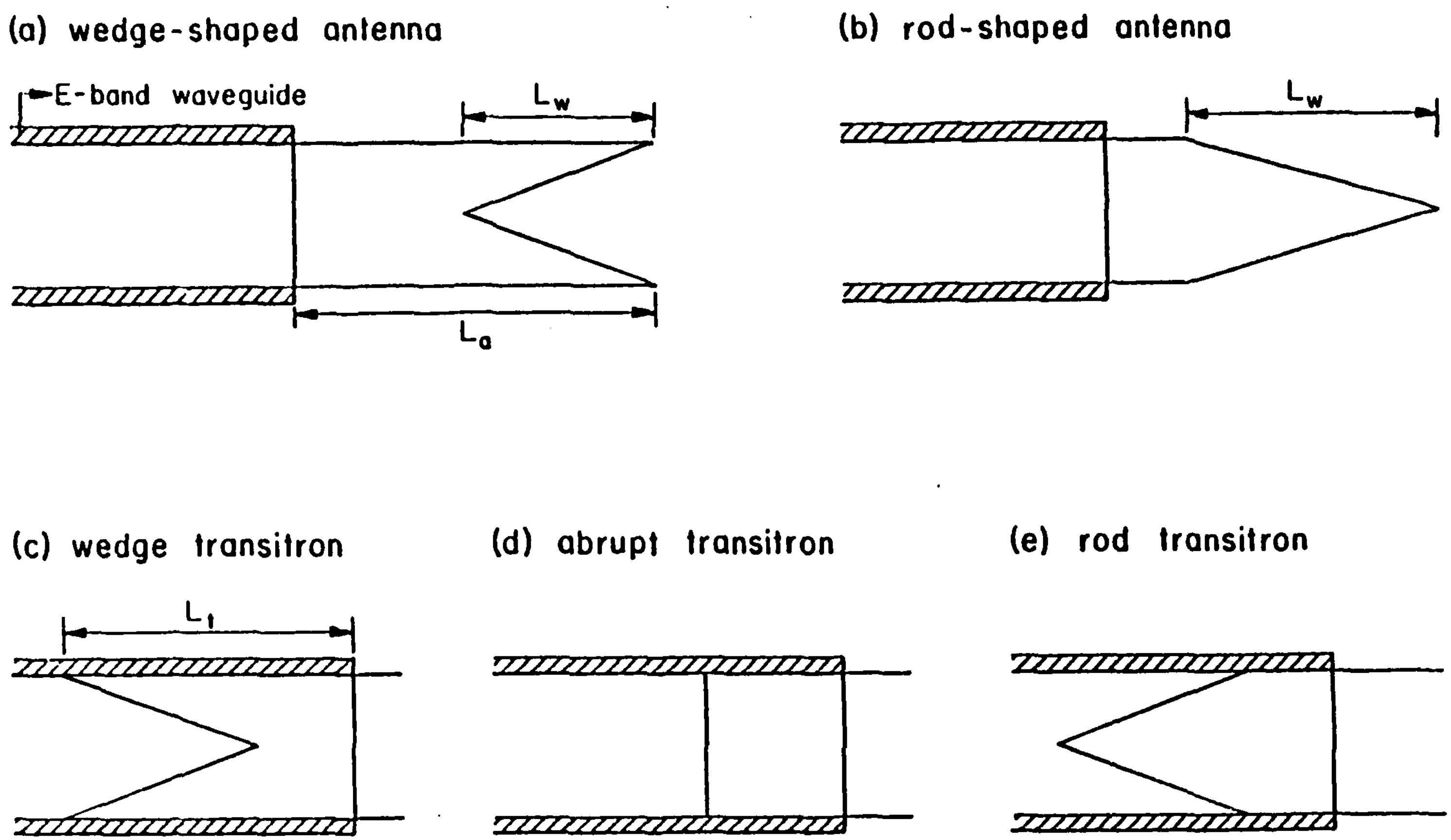


Figure 5. H-plane view.

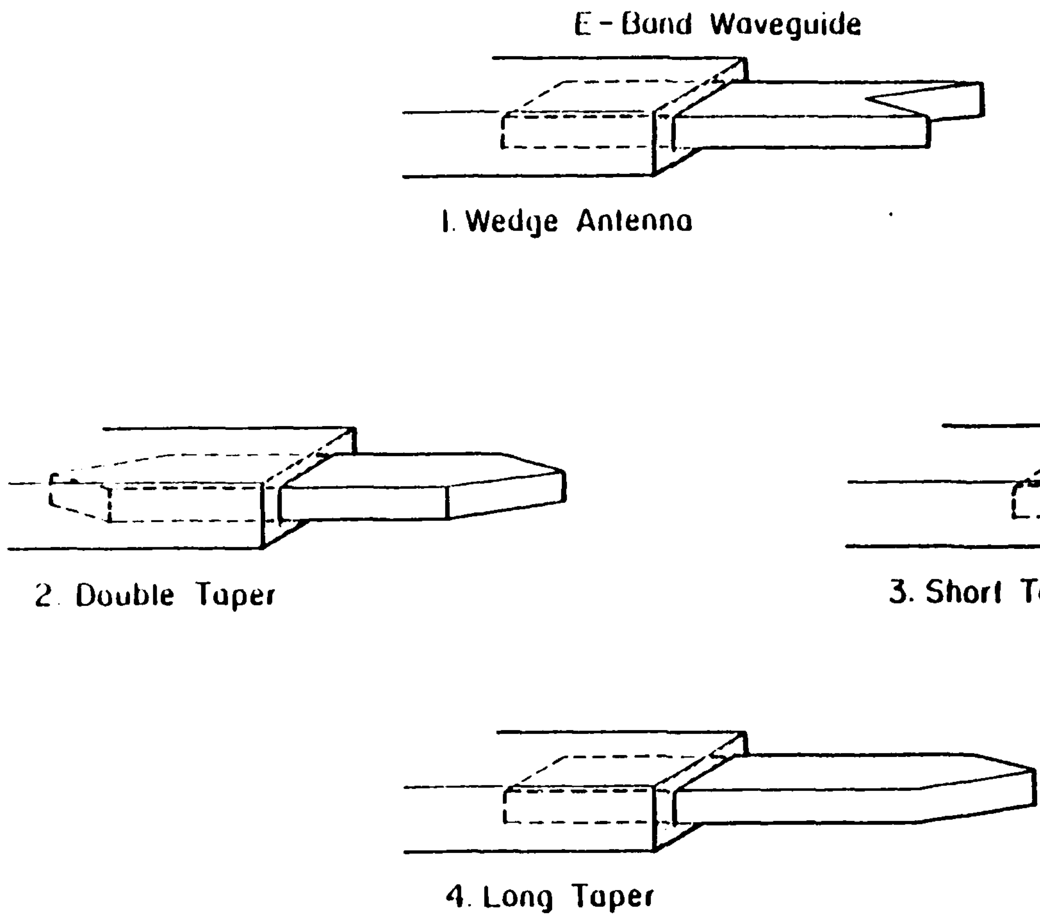


Figure 6. Test antenna shapes.

tested. They also find from experimentation that this antenna has extraordinary image scanning capabilities due to its relatively high resolution. Another discovery is that the transition end is not a crucial factor in the overall performance of the antennas, which is consistent with the findings of Chang and Mittra.

In summary, the design factors to be considered in the fabrication of the dielectric rod antenna are the length and shape of the transition end, length of the rod inside the waveguide, and the length and shape of the radiating end. The least crucial are the length and shape of the rod inside the waveguide.

III. PERIODICALLY MODULATED SURFACE-WAVE ANTENNAS

If a radiation pattern is desired in some direction other than endfire, the dielectric rod antenna must be replaced by another type of antenna. Kobayashi, Paleta, and Ray [10] all discuss the periodically modulated surface-wave antenna, also called the leaky-wave antenna, as the solution to this problem. This antenna is simply a dielectric rod antenna with perturbations which serve as radiators of power. These perturbations along the rod form an array which is frequency steerable and may be excited by a travelling wave along the surface of the antenna. If the effects of the perturbations at feed and terminal ends are neglected, along with the effects of mutual coupling between elements and the backward waves reflected at discontinuities, the pattern may be described by the array factor

$$AF = A_o \sum_{k=0}^{N-1} Z^k = A_o \frac{1-Z^N}{1-Z}$$

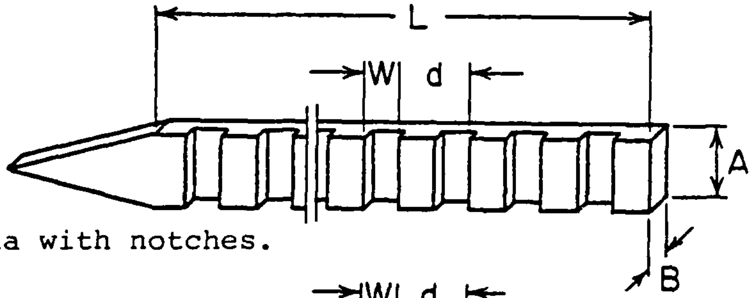
where A_o is a constant, N is the number of radiating elements, and Z is a complex number given by $Z = \alpha e^{j\psi}$ where $0<\alpha<1$ is the amplitude ratio between two adjacent elements, i.e., the attenuation constant of the elements.

$\psi = k_o d(\cos\theta - \beta_z / k_o)$ is the phase difference between adjacent elements, where k_o and β_z are, as previously defined, the free-space wave number and the dielectric phase constant in the z-direction, respectively.

Figure 7 gives several examples of this type of antenna, along with the coordinate system used in the analysis. In order to precisely predict the radiation pattern of a periodically modulated surface-wave antenna, the type and dimensions of each perturbation must be known. To produce the pattern properly, the array factor is multiplied by the space factor of each perturbation and by the element pattern of an infinitesimal dipole. Therefore, the desired pattern is directly dependent on the type of perturbation chosen. Kobayashi investigates several types of perturbations, beginning with an antenna with notches (Figures 7 (a,b)). This type of antenna has a very high level of unnecessary lobes in the endfire direction, which is mostly due to the residual surface-wave energy. Plating the inside walls with silver (Figure 7 (c)) suppresses the residual lobes below the main beam level, but not to a satisfactory level. Another configuration tested, the antenna with silver strips (Figure 7 (d)), shows better performance in that it has a higher main beam and the endfire radiation is no longer attributable to residual surface waves. It is also possible to place strips on the narrower side of the dielectric rod (Figure 7 (e)), but refers to a different radiation pattern than that produced by the antenna with strips on the broad side. It also has a complicated polarization characteristic. In general, the leaky-wave antenna tends to have high side-lobe levels.

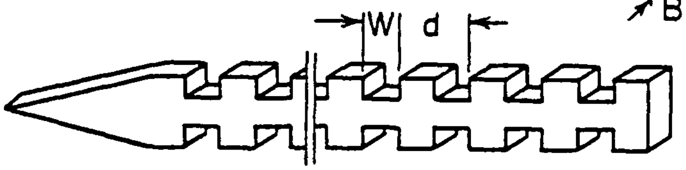
Despite the high side-lobe level of the configuration in Figure 7 (d), it tends to perform in the manner desired. If the high sidelobes are diminished, the antenna is useful. Also, the radiation from the feed and terminal points must be minimized so that they do not interfere with the radiation from the metal strips. Both

(a)



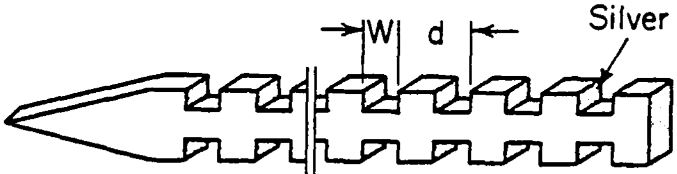
Antenna with notches.

(b)



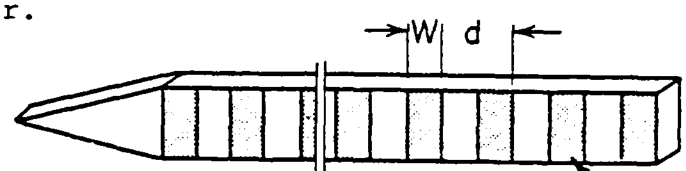
Antenna with notches.

(c)



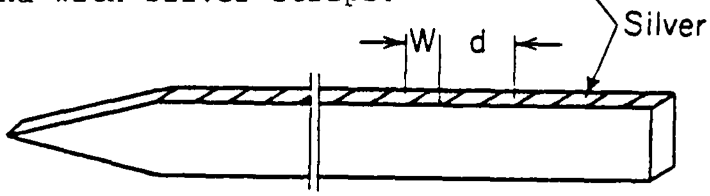
Same as (b) except with inside walls plated with silver.

(d)



Antenna with silver strips.

(e)



Antenna with silver strips placed on narrower side.

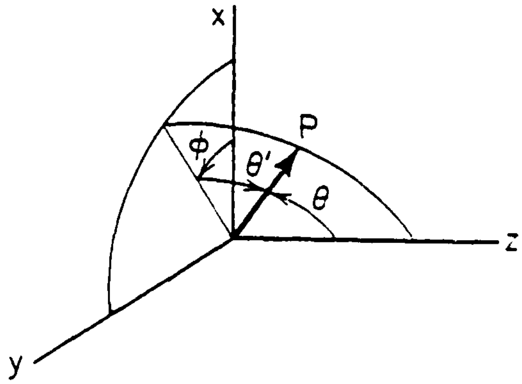


Figure 7. Examples of periodically modulated surface-wave antennas and the coordinate system.

Paleta and Ray investigate this situation. The transition from the feed point to the perturbed section of the antenna should be gradual since an abrupt beginning of the radiating portion of the antenna will create a sizable discontinuity and cause serious pattern degradation due to the extensive radiation which occurs in the vicinity of the discontinuity. Undesirable radiation from the antenna termination can be minimized by appropriately choosing the length of the structure. Since only a small amount of power is radiated by each individual strip perturbation, the number of strips must be chosen so that a negligible amount of power remains at the termination. Since narrow strips produce very little radiation while wide strips are very efficient radiators, the obvious solution to the problem of excessive endfire radiation at the terminal end is strip tapering, i.e., placing narrow strips near the feed and gradually increasing the strip width, while holding the strip spacing constant, as the amount of power travelling down the rod decreases. Paleta and Ray both try this method, each obtaining better results than those obtained for antennas with constant strip width. Paleta investigates several linear taper distributions, since the decrease in power is, for the most part, linear. He finds that the distribution,

$$w_n = \begin{cases} 0.3+n\,(0.025)mm & 0<n \leqslant 45 \\ 1.43mm & n>45 \end{cases}$$

provides the best results at 81.5 GHz. It has both a sharper main beam and lower side-lobe levels. Ray obtained similar results.

Paleta continues further investigation of different types of perturbations, such as tapering both length and width, circular perturbations, and rectangular perturbations, i.e., thin strips placed parallel to the guiding structure rather than perpendicular to it. None of these modifications are able to produce results comparable to that of tapered strip width antennas.

All of the antennas considered have the same basic problems. These are 1) two main beams of almost equal amplitude and 2) very narrow beams in the E-plane but wide beams in the H-plane. Figure 8 illustrates both of these characteristics. In order to solve the first problem, a method needs to be devised such that radiation from one beam interferes constructively with the radiation from the other beam, hopefully doubling the radiation of a single beam (3 dB increase). Several approaches have been made to this problem, including slot waveguide arrays constructed from metal waveguides, and metallized dielectric waveguides. There are several drawbacks from these configurations. For instance, for the slot waveguide, the wall thickness of the waveguide is on the order of one wavelength and must be machined down to very narrow dimensions before the radiating slots can be cut. In the case of the metallized dielectric waveguide, it is extremely difficult to feed, the desired results are not easily repeatable, and the expected 3 dB increase is not obtained.

One configuration that is simpler to construct and more reliable than the others is the image line antenna, in which a reflective surface (ground plane) is placed behind the antenna structure. This causes the backward beam to be reflected forward so that it adds to the main forward beam. The reflector must be placed so that the reflected power adds in phase with the forward radiated power. Results show that the side-lobe level in the E-plane is negligible, but the H-plane exhibits an extremely high side-lobe level. Itoh and Adelseck [11], Solbach and Adelseck [12], Solbach and Wolff [13], Bahl and Bhartia [14], and Schwering and Peng [15] all investigate this type of antenna, making use of modifications such as a trapped image guide [11], dielectric gratings for the perturbations [15], and artificial dielectric materials [14]. The latter paper also addresses another problem common to the image guide antenna, its inability to frequency scan. As the operating frequency is varied, the reflected and forward beams no longer add in phase and the

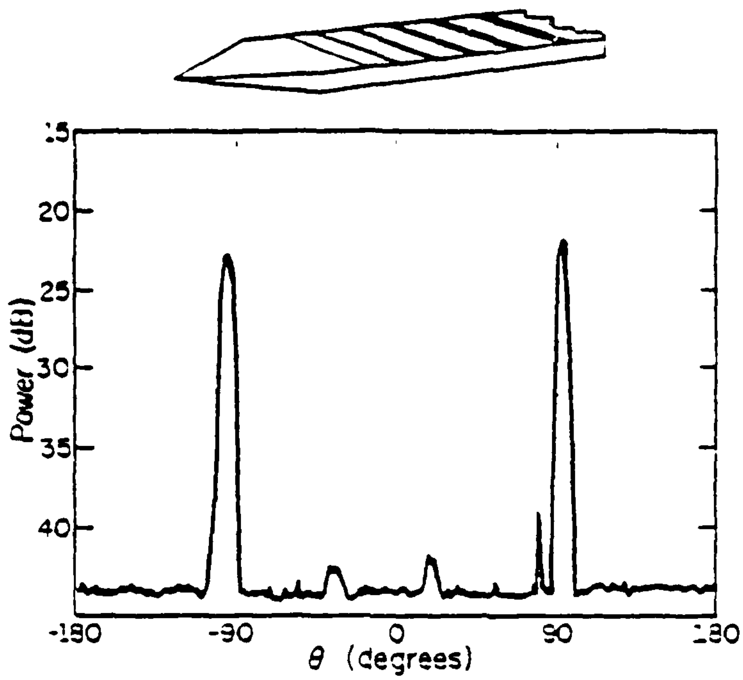


Figure 8a. E-plane pattern of a tapered strip width antenna.

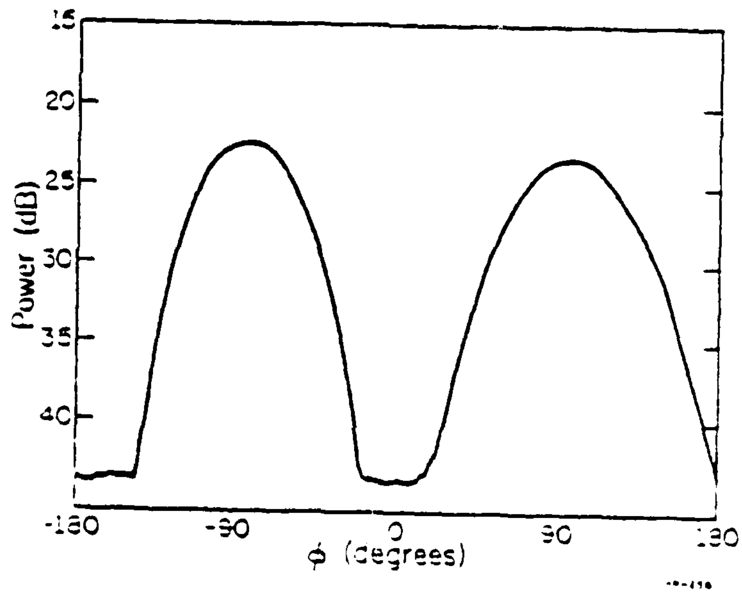


Figure 8b. H-plane pattern of a tapered strip width antenna.

directive gain quickly deteriorates. Scanning is achieved as a result of the frequency dependence of the complex propagation constant. In the artificial dielectric antenna, the complex propagation constant can be changed by changing the frequency or by changing the dielectric constant of the medium. In another configuration, Seiler and Mathena [16] use "diffractions electronics" to control the direction of radiation.

Trinh [17] et al. try to resolve the second basic problem of periodically modulated surface-wave antennas, the wide beams in the H-plane. In an effort to minimize the H-plane beamwidth, a trough was cut along the entire length of the ground plane and the antenna mounted below the surface of the ground plane, in the hopes that the walls of the ground plane could be used to focus the power in the H-plane. This structure still produces beams in the H-plane which are unsatisfactorily wide. The addition of a metal flair (see Figure 9) greatly reduces the H-plane beamwidth. By proper adjustment of the horn flair, the H-plane beamwidth may be decreased, and the directive gain increased. The optimum flair angle α is experimentally determined, and the maximum directive gain decreases for angles both larger and smaller than this optimum value. It has also been found that the larger the flair slant length, the greater the directive gain.

To summarize, the periodically modulated surface-wave antenna demonstrates relatively good performance, despite the fact that it is plagued with high side-lobe levels and a wide H-plane pattern. Through modifications of the basic configuration, these problems may be alleviated. The antenna may be designed to radiate in any direction from -180 to +180 degrees, which differs from the dielectric rod antenna which is restricted to the endfire direction.

IV. COMPLEX PROPAGATION CONSTANT

Up to this point, only the fields outside of the dielectric have been discussed.

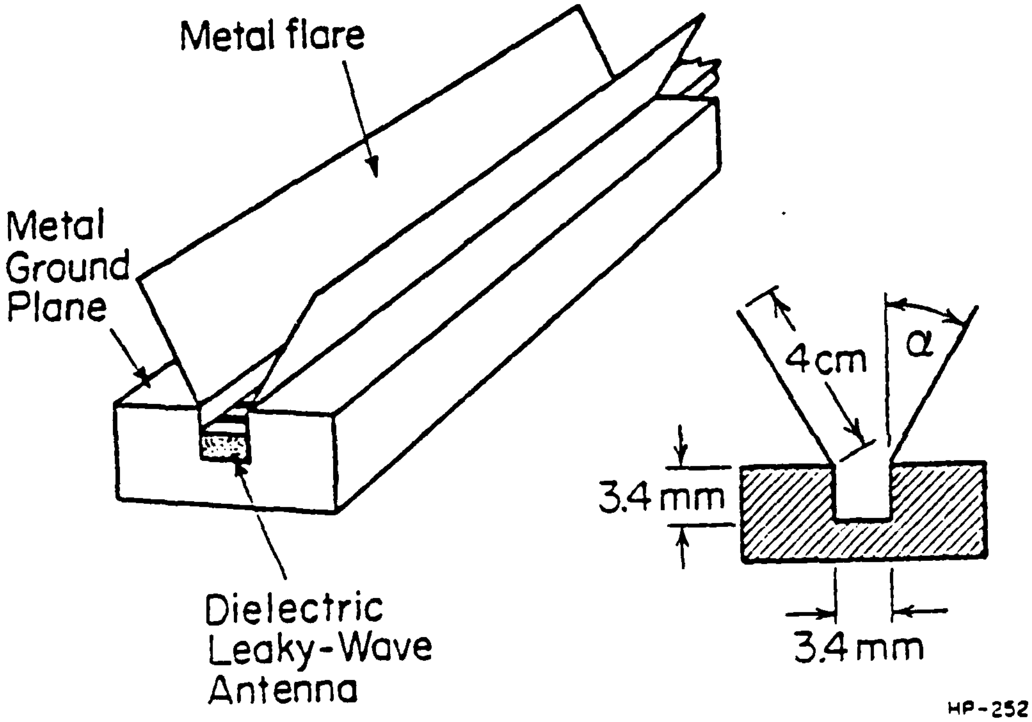


Figure 9. Horn-image guide antenna structure.

Since the basis of these antennas is the dielectric rod antenna, there should be an investigation into the behavior of the fields inside the dielectric. It is most important to determine the complex propagation inside the dielectric, $k_z = \beta_z - j\alpha_z$. This is of great importance since β_z determines the direction of the main beam, and α_z determines the beamwidth. This problem is approached in several ways. McLevige uses Maxwell's equations and boundary conditions in this problem which becomes so complicated that it ultimately requires the use of a digital computer for a numerical solution. Ray experimentally determines the propagation constant with two different approaches, near-field probing and far-field inference. Near-field probing is performed on the nonmetallized side of the leaky-wave antenna, in order to avoid the drastic field variation which arises as a result of the boundary condition that $\vec{E} = \vec{0}$ on the strips. The distance between maxima (minima) is $\lambda_g/2$. β_z may therefore be determined by $\beta_z = 2\pi/\lambda_g$. α_z is determined from $\alpha_z(np/mm) = (\Delta/\lambda_g)\ \ln 10$, where $\Delta \equiv$ difference in height of adjacent peaks. In far-field inference, β_z is determined by observing that $\beta_z = k_o(\sin\theta' - n\lambda_o/d)$ where n = -1 from design considerations and d is the spacing of the perturbations ([1] Appendix). α_z is determined from the relation $\left|AF(\theta_{3dB})\right| = \frac{1}{\sqrt{2}}\left|AF\right|_{MAX}$ which is solved for the α_z that gives the measured beamwidth.

If the dielectric guiding structure is comprised of several materials with different dielectric constants, we may apply the method of effective dielectric constants (EDC) to simplify the analysis. Schwering and Peng, Ray, Rudokas, and Yang all discuss this method. Figure 10 shows a guiding structure comprised of more than one dielectric material. If each region, alone, is extended to $\pm\infty$ in the x-direction, each may be considered as a simple multilayered slab waveguide. By

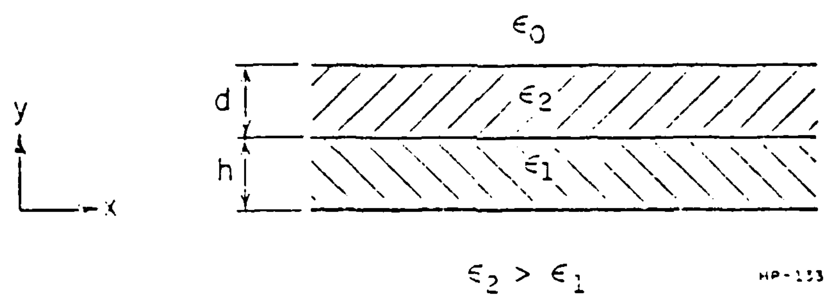


Figure 10. Guiding structure comprised of more than one dielectric material.

phase matching at each interface, the propagation constant, in the y-direction, may be determined for each region. Each region may then be modeled by a homogeneous region with a dielectric constant chosen such that the propagation constant is identical to that of the original structure.

Mittra and Kastner [18] use another method for determining the complex propagation constant in the leaky-wave structure. This method is based on the spectral domain approach that formulates the eigenvalue problem in the Fourier transform domain by replacing the dielectric rod by a slab using the EDC method, and completing the analysis for the new geometry.

All of the approaches mentioned above tend to yield satisfactory results. It is up to the designer to decide which method is simplest for the case being considered.

V. COMPONENTS FOR MILLIMETER-WAVE INTEGRATED CIRCUITS

In addition to the fact that it is lightweight and compact, and the performance capabilities demonstrated by it, the dielectric antenna is of interest due to its ability to be integrated with other components into a millimeter-wave network. At microwave frequencies, metal waveguides and coaxial cables are used for transmission of the wave. At millimeter-wave frequencies, the physical dimensions of these waveguides become quite small and rather lossy. They are also expensive to manufacture at millimeter-wave frequencies. Dielectric waveguides are an alternative to this problem. Figure 11 gives the basic geometries for dielectric waveguides for integrated circuitry. Unlike optical fibers, millimeter-wave waveguides are usually rectangular and are often mounted on a metallic ground plane for planar system integration. The cross-sectional dimensions of a dielectric

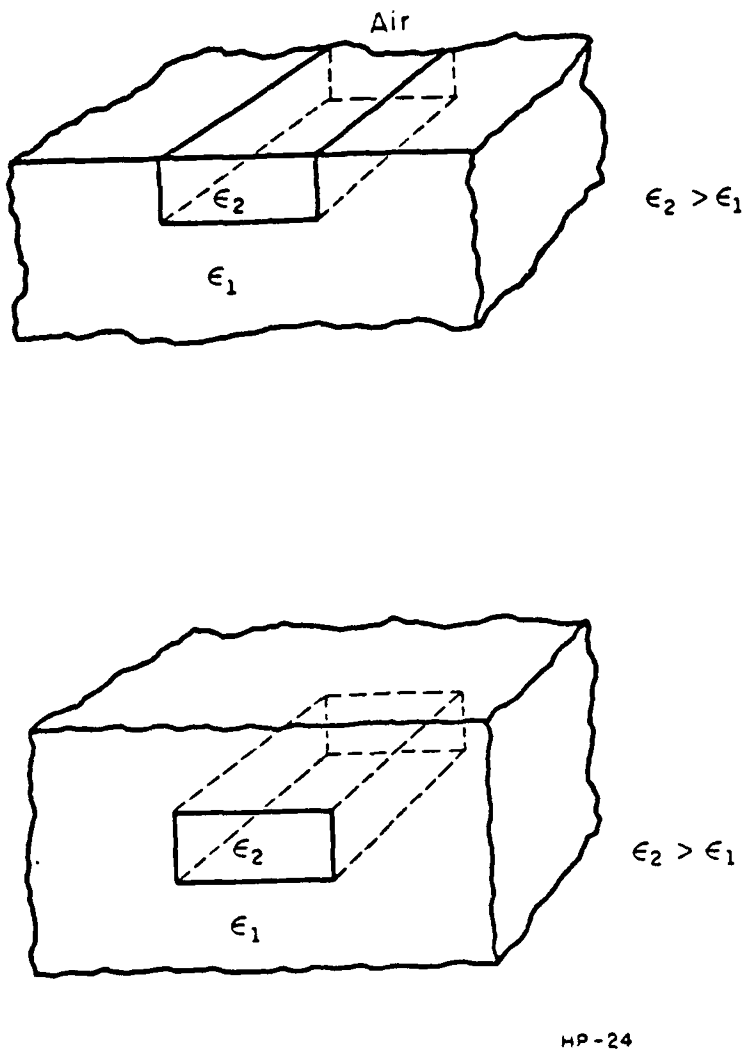


Figure 11. Basic geometries of dielectric waveguides for integrated optical circuitry.

guide are required to be on the order of the wavelength in an unbounded dielectric for single-moded operation, and are almost always designed to be single-moded. A strip of dielectric is immersed in another dielectric of smaller permittivity so that total internal reflection occurs at the dielectric interfaces, thus confining the wave to propagate down the guide in the region of the highest dielectric constant. Figure 12 shows other dielectric waveguiding structures. By getting maximum usage out of the dielectric waveguides, it is feasible to produce monolithic integrated structures which place a complete transmitter or receiver on a single dielectric slab. Components such as oscillators, mixers, phase shifters, and directional couplers can be molded into a single sheet of dielectric material. Integrated circuits may therefore be easily and inexpensively produced by injection molding or stamping.

McLevige examines several types of dielectric waveguides. Beginning with Maxwell's equations, he determines the fields in the waveguide. The complexity of the analysis requires computer analysis, and the computer programs used are included in the appendix of the paper. Yang performs a similar analysis. Both support their theoretical findings with experimental results.

Rudokas and Bhooshan examine more complicated geometries. They both present and analyze components used in millimeter-wave integrated circuits. Some examples of the devices are the inverted stripguide (Figure 13), the distributed directional coupler (Figure 14), and the ring resonator (Figure 15). Bhooshan also presents results of experiments conducted to determine the effectiveness of the inverted stripguide in the construction of components for millimeter-wave integrated circuits.

The results of the experiments show that the inverted stripguide and the distributed directional coupler demonstrate characteristics excellent for integrated

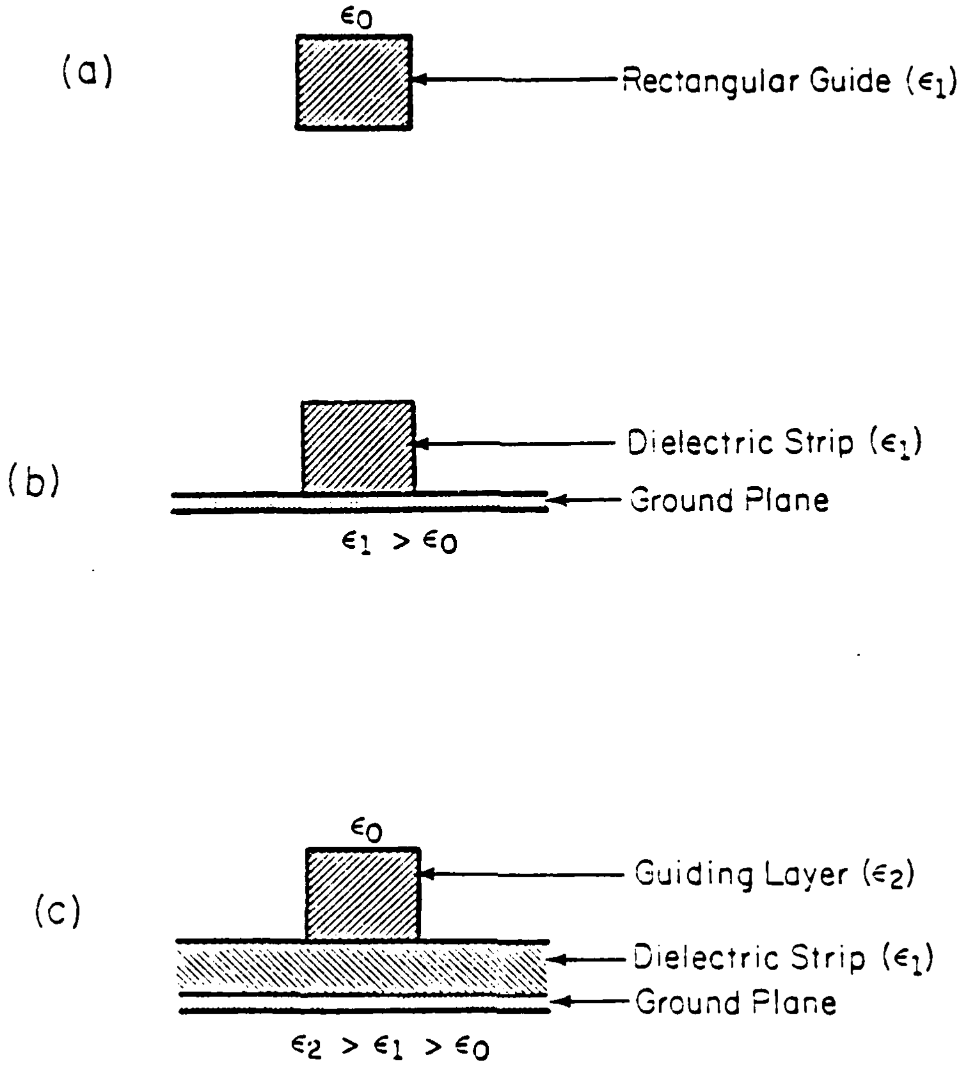


Figure 12. Dielectric waveguiding structures
a. suspended guide
b. image guide
c. insulated strip guide

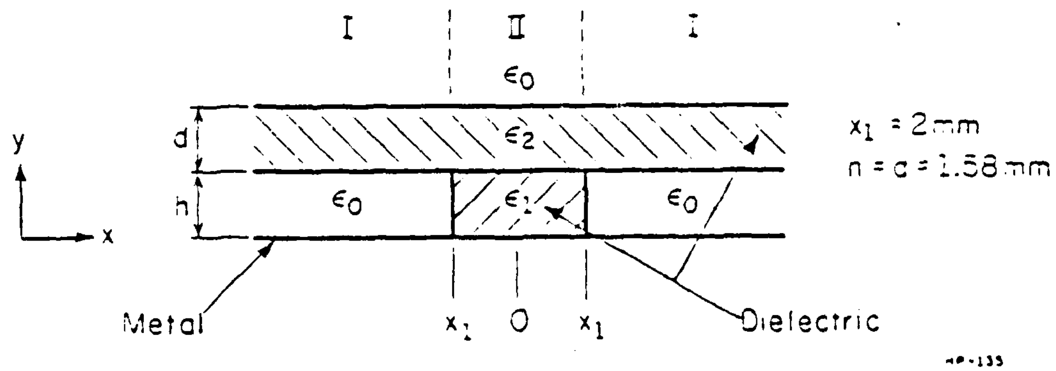


Figure 13. Inverted strip guide.

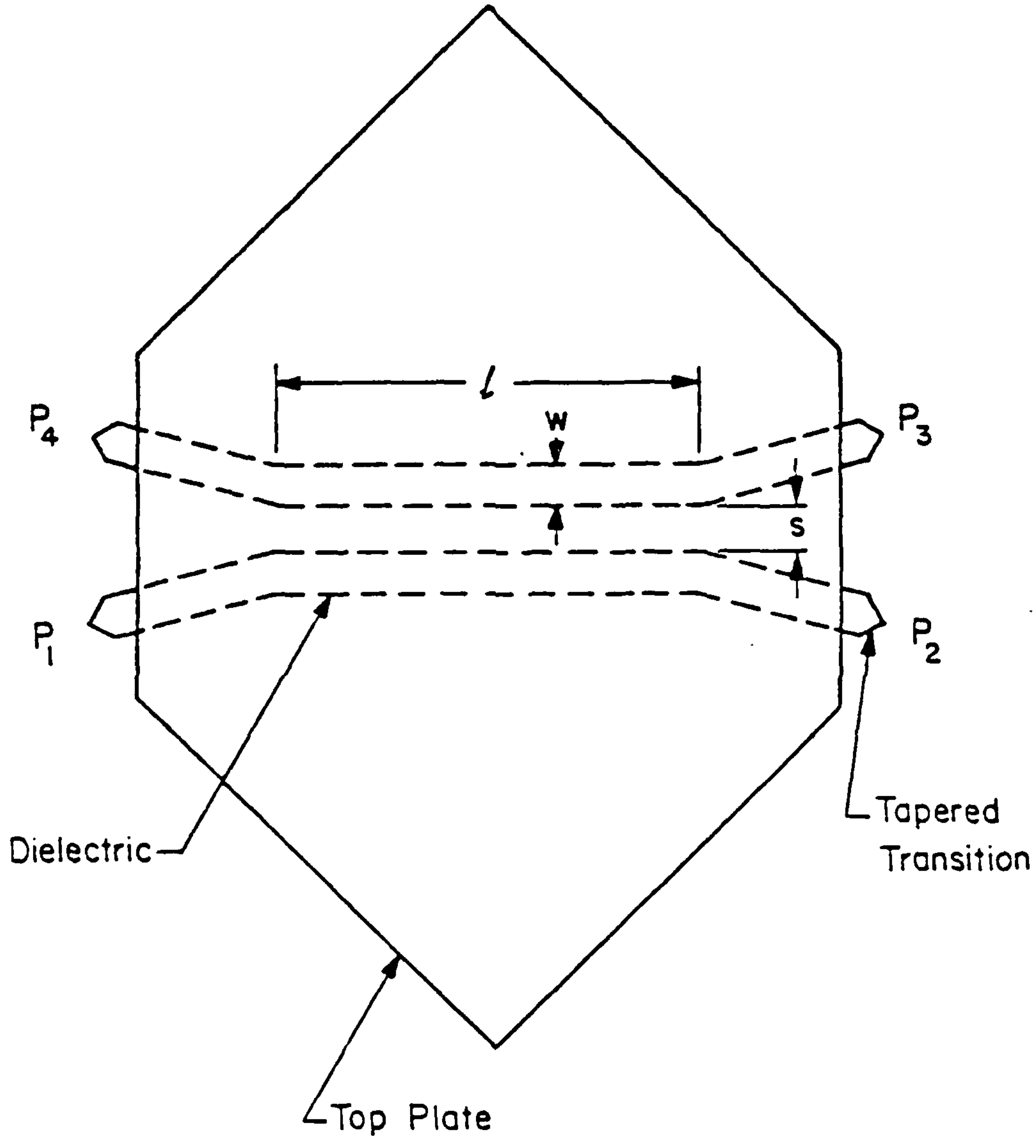


Figure 14. Top view of the directional coupler with the four connecting guides.

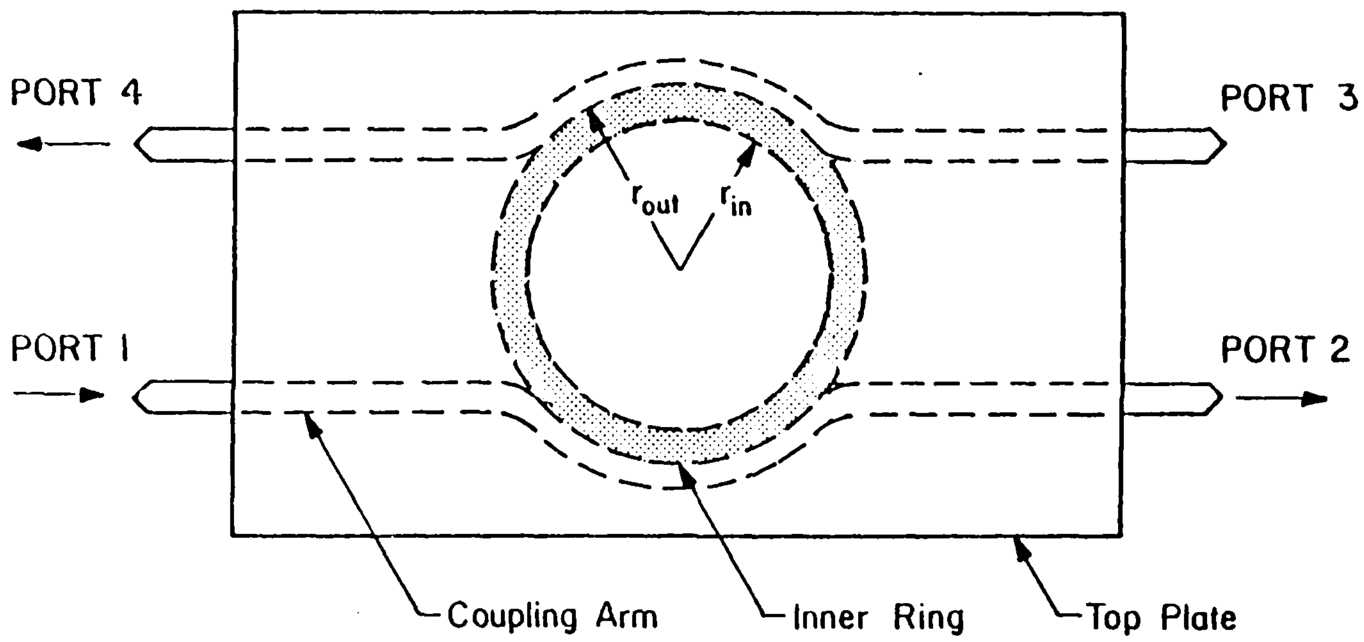


Figure 15. Ring resonator.

circuit applications, while the ring resonator does not perform very efficiently in this type of network.

VI. CONCLUSION

The literature reviewed in this report illustrates the versatility of the dielectric antenna and associated integrated circuit components. There are numerous articles written on this topic, concerning variations of the basic configuration and other ways of improving the performance of the antenna. The articles presented here only begin to scratch the surface, but they indicate the starting point and important factors which must be considered in the design of an antenna of this type.

VII. BIBLIOGRAPHY

[1] S. Kobayashi, "Dielectric Antennas for Millimeter-Wave Applications," M.S. Thesis, University of Illinois at Urbana-Champaign, 1980.

[2] R. J. Paleta, "A Study of Dielectric Rod Leaky-Wave Antennas," M.S. Thesis, University of Illinois at Urbana-Champaign, 1981.

[3] W. V. McLevige, "Analysis of Dielectric Waveguides for Millimeter-Waves and Optical Integrated Circuits," M.S. Thesis, University of Illinois at Urbana-Champaign, 1975.

[4] P. Yang, "A New Method for the Analysis of Dielectric Waveguides for Millimeter-Wave and Optical Integrated Circuits," M.S. Thesis, University of Illinois at Urbana-Champaign, 1978.

[5] R. S. Rudokas, "Analysis of Inverted Strip Dielectric Waveguides and Passive Devices for Millimeter-Waves," M.S. Thesis, University of Illinois at Urbana-Champaign, 1977.

[6] S. V. Bhooshan, "Multimode Waveguide Components for Millimeter-Wave Integrated Circuits," M.S. Thesis, University of Illinois at Urbana-Champaign, 1979.

[7] B. H. Ahn, "Measurement of the Indices of Refraction and the Absorption Coefficients of Dielectric Materials in the Millimeter-Wave Region," *Journal of Applied Physics*, vol. 54, no. 4, pp. 2123-2124, April 1983.

[8] A. Chang and R. Mittra, "Experimental Study of Dielectric Antenna at the Frequency Range of 76-80 GHz," Interim Technical Report, Electromagnetics Laboratory, University of Illinois at Urbana-Champaign, October 1983.

[9] S. H. Doran and R. Mittra, "An Experimental Study of Dielectric Rod Antenna for Millimeter-Wave Imaging Applications," Technical Report, Electromagnetic Communication Laboratory, University of Illinois at Urbana-Champaign, March 1985.

[10] S. L. Ray "Dielectric Leaky-Wave Antennas for Millimeter-Wave Applications," M.S. Thesis, University of Illinois at Urbana-Champaign, 1981.

[11] T. Itoh and B. Adelseck, "Trapped Image Guide Leaky-Wave Antennas for Millimeter-Wave Applications," *IEEE Transactions on Antennas and Propagation*, vol. AP-30, no. 3, pp. 505-509, May 1982.

[12] K. Solbach and B. Adelseck, "Dielectric Image Line Leaky-Wave Antenna for Broadside Radiation," *Electronics Letters*, vol. 19, no. 16, pp. 640-641, 4 August 1983.

[13] K. Solbach and Ingo Wolff, "Dielectric Image Line Groove Antennas for Millimeter-Waves," *IEEE Transactions on Antennas and Propagation*, vol. AP-33, no. 7, pp. 690-706, July 1985.

[14] I. J. Bahl and P. Bhartia, "Leaky-Wave Antennas Using Artificial Dielectrics at Millimeter-Wave Frequencies," *IEEE Transactions on Microwave Theory and Techniques*, vol. MTT-28, no. 11, pp. 1025-1212, November 1980.

[15] F. K. Schwering and S. T. Peng, "Design of Dielectric Grating Antennas for Millimeter-Wave Applications," *IEEE Transactions on Microwave Theory and Techniques*, vol. MTT-31, no. 2, pp. 199-209, February 1983.

[16] M. R. Seiler and B. M. Mathena, "Millimeter-Wave Beam Steering Using 'Diffraction Electronics'," *IEEE Transactions on Antennas and Propagation*, vol. AP-32, no. 9, pp. 987-990, September 1984.

[17] T. N. Trinh, R. Mittra, and R. J. Paleta Jr., "Horn Image-Guide Leaky-Wave Antennas," *IEEE Transactions on Microwave Theory and Techniques*, vol. MTT-29, no. 12, pp. 1310-1314, December 1981.

[18] R. Mittra and R. Kastner, "A Spectral Domain Approach for Computing the Radiation Characteristics of a Leaky-Wave Antenna for Millimeter-Waves," *IEEE Transactions on Antennas and Propagation*, vol. AP-29, no. 4, pp. 652-654, July 1981.

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